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A Generalized Parameter Tuning Method of Proportional-Resonant Controllers for Dynamic Voltage Restorers

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ABSTRACT

Temporary voltage swells and sags appear with high frequency in electric power systems, and they significantly affect sensitive loads such as industrial manufacturing or communication devices. This paper presents a strategy to design proportional-resonant controllers for three full-bridge voltage-source converters with a common DC-link in dynamic voltage restorer systems. The proposed controllers allow the system to quickly overcome temporary unbalanced voltage sags. Simulation results carried out in MATLAB/Simulink and experimental results implemented in a Typhoon HIL402 device demonstrate the ability of the proposed design method. The results show that the system with the proposed controllers can ride-through single-phase or double-phase voltage sags up to 55% and three-phase voltage sags up to 70% in a duration less than one grid-voltage cycle.

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1. INTRODUCTION

Power quality disturbances such as voltage sag and swell have been causing serious concerns for modern distribution systems operated at low- and medium-voltage levels. The percentage of voltage sags caused by single line-to-ground fault, double line-to-ground fault, and balanced three phase-to-ground fault in power systems are 68%, 19%, 13%, respectively [1]. Such voltage variations in short durations (less than 60 seconds) lead to improper operation of sensitive loads, while longer voltage variations can result in sustained interruptions or failures. Therefore, mitigating voltage sags and swells in low- and medium-voltage distribution systems is critical.

One of the most solution to improve voltage regulation is dynamic voltage restorers (DVRs). The operating principle of DVRs is to inject appropriate voltage in series and synchronism with the distorted AC grid source to compensate for the amount of voltage sag or swell [2]-[4]. A DVR system includes an energy storage, a three-phase voltage-source inverter, and series connected transformers between an AC grid source and a load. Regarding the three-phase voltage-source inverters, it is necessary to control the positive-, negative-, and zero-sequence components to compensate the voltage sag or swell of each phase [5]. Therefore, the control scheme is complicated and insufficiently reliable. To eliminate voltage sags and swells, several alternative topologies of DVR have been introduced such as a three-phase inverters with a neutral point created by a DC-link capacitor, three-phase four-wire inverters, and three single-phase full-bridge inverters with a common DC-link capacitor [6],[7]. The latter is preferred because of the simple pulse-width modulation (PWM) method.

In DVR systems using three full-bridge inverters with a common DC-link capacitor, several control approaches have been suggested. A coordination of both feed-back and feed-forward control in

synchronously rotating frame *dq* is introduced in [8],[9]. However, this topology does not have current loops to compensate for the losses in the transformers and output filter, which leads to a slow voltage response. Another approach is to implement proportional-resonant (PR) or H_{inf} control for the voltage loop and proportional (P) control for the current loop in the stationary frame $\alpha\beta$ [10]. However, controlling only two components of $\alpha\beta$ frame as presented in this approach is not effective when the voltage sag is unbalanced, which is the typical case in practice. A P control for both voltage and current loops with independent controllers between each phase is proposed in [11]. This approach is unable to eliminate the steady-state error in spite of its fast response. A multi-loop with PI controllers is applied to DVR systems in [12], but steady-state error also exists.

From this literature review, the control strategy using PR controllers for both current and voltage loops is the most effective for DVR systems to improve dynamic response and eliminate steady-state error [13]. To the best knowledge of the authors, little research has been done to elaborate on the process of determining parameters for PR controllers, which is highly complicated. Several studies apply trial and error procedures to obtain the parameters of PR controllers [14]. Another approach to design PR controllers is based on the SISO design tool in MATLAB and system dynamic response [15]. Such methods are time-consuming and not generalized.

This paper proposes a systematic and generalized design method for PR voltage and current controllers of three single-phase full-bridge inverters with a common DC-link capacitor in DVR systems. The major contributions of this paper includes:

a) An equivalent circuit of full-bridge inverters and series connected transformers in a DVR system. Unlike other existing methods, the model of the series transformers is taken into account in this paper when designing the controllers for the DVR system.
b) A method to design parameters of PR controllers in the frequency domain for both current and voltage loops to guarantees system stability with a desired cross-over frequency and phase margin. The discretization in the *z*-domain of the designed PR controllers for digital implementation is also included.
c) An approach to verify the proposed controller by hardware-in-the-loop (HIL) real-time experiments using a Typhoon HIL402 device. HIL simulation has been highly recommended as an effective design approach with the ease in modifying controller parameters and creating different operating scenarios of grid voltage [16]-[18].

The remainder of this paper is organized as follows. Section 2 describes the general control topology and the model of a DVR system including the equivalent circuit of full-bridge inverters and series transformers. The proposed design in the frequency domain of the PR current and voltage controllers is presented based on the developed equivalent model. The discretization of the designed PR controller is also included. Section 3 demonstrates the efficacy of the proposed method by off-line simulation and HIL real-time experiments using Typhoon HIL402 system.

2. CONTROL SCHEME

A scalar control scheme for three full-bridge inverters with a common DC-link capacitor in a DVR system is shown in Figure 1. Similar controllers are applied separately for each phase corresponding to each H-bridge inverter. In this control scheme, voltages and currents of all phases are controlled independently using nested control loops. At each phase, the outer voltage control loop regulates the voltage at the secondary side of the transformer while the inner current loop regulates the output current of the H-bridge inverter. The set points of the current loop are the output of the voltage loop, while the set points of the voltage loop are calculated based on the root-mean-square (RMS) values of the desired voltage (for example, 220 V) and the grid voltage. In the proposed control approach, since the instantaneous voltage and current are measured, PR controllers are chosen to eliminate the steady-state error. The resonant frequencies of the PR controllers are equal to the grid frequency. In addition, phase-locked loop (PLL), which is required for grid synchronization, is implemented by measuring the voltage at each phase of the grid source. In this paper, a phase-locked loop (PLL) algorithm based on a second-order generalized integrator phase-locked loop (SOGI PLL) is used [19].

To properly design PR controllers, the model of series transformers in a DVR system should be taken into account. The equivalent circuit of each H-bridge and the simplified model of the transformer referred to the secondary side are shown in Figure 2. The leakage impedance of the transformer and capacitance result in a second-order low-pass filter. The equivalent impedance referred to the secondary side of the transformer is given as follows:

$$Z_{eqS} = \left(\frac{r_p}{N^2} + r_s\right) + j\left(\frac{x_{s\,p}}{N^2} + x_{s\,s}\right) = R + j\omega_1 L_s \,, \tag{1}$$

where N is the turns ratio of the transformer, r_p and r_s are the resistances of the primary and secondary windings, $x_{s\,p}$ and $x_{s\,s}$ are the leakage inductances of the primary and secondary windings, and R and L_s are the resulting resistance and inductance of the equivalent impedance.

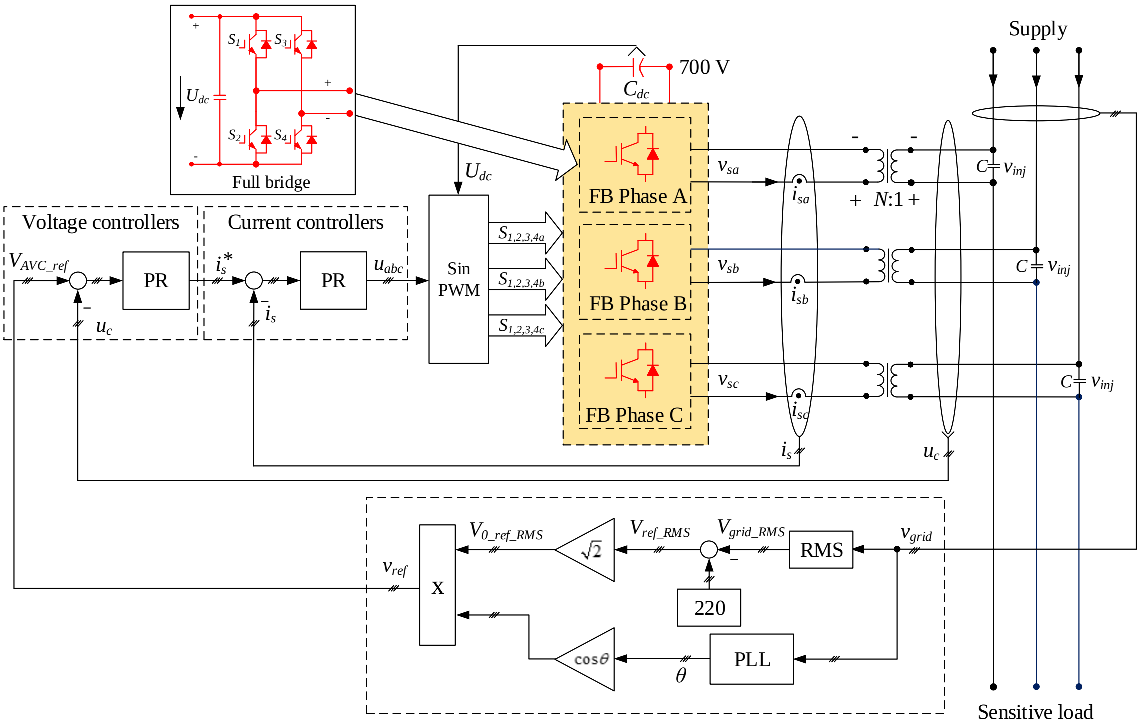


Figure 1. Control topology of the inverter in DVR

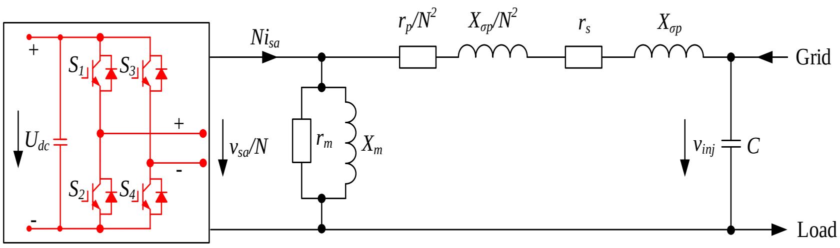


Figure 2. Equivalent circuit of per phase and series connected transformer

A PR controller has an infinite gain at a selected resonant frequency; thus, the zero steady-state error or the harmonic at this frequency can be eliminated. The transfer function of a PR controller is mathematically expressed as follows [20]:

$$G_{PR}^{c}(s) = k_p + \frac{k_r s}{s^2 + (\omega_1)^2}\,. \tag{2}$$

where k_p and k_r are the coefficients of the PR controller while ω_1 is a selected frequency. The frequency response characteristics of the PR controller are calculated as follows:

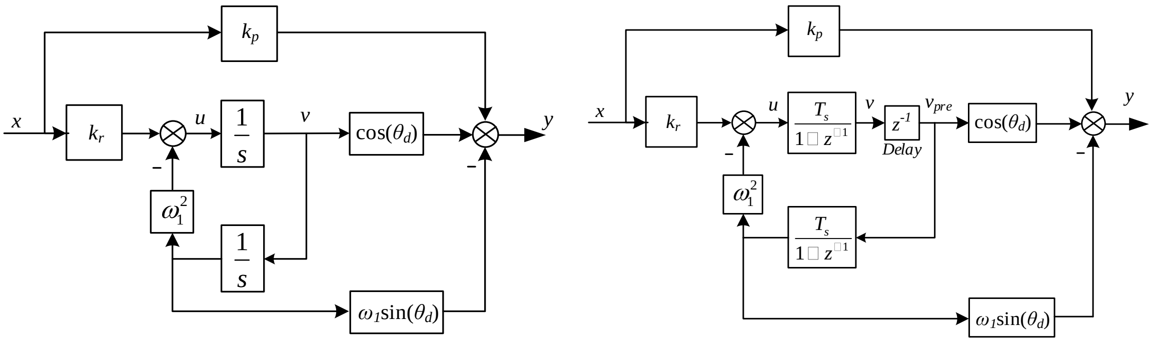
$$\left|G_{PR}(j\omega)\right| = \frac{\sqrt{k_p^2\left(\omega_1^2 - \omega^2\right)^2 + k_r^2\omega^2}}{\left(\omega_1^2 - \omega^2\right)} \quad (3)$$

$$\angle G_{PR}(j\omega) = \arctan\left[\frac{k_r\omega}{k_p\left(\omega_0^2 - \omega^2\right)}\right] \quad (4)$$

From (3), it can be seen that the PR controller has an infinite gain at ω_1. In [21], the transfer function of the PR controller with the delay compensation can be written as follows:

$$G_{PR}^{c}(s) = k_p + k_r\frac{\cos(\theta_d)s - \omega_1\sin(\theta_d)}{s^2 + (\omega_1)^2}. \quad (5)$$

The delay can be compensated by adding a lead angle $\theta_d = \omega_1(kT_s)$ to the inverse Park transform angle, where k is an integer representing the number of periods to be compensated [21]. In this paper, k is set to 1. To avoid an algebraic loop during the discrete implementation, it is suggested that the direct integrator is discretized using the forward method while the feedback integrator is discretized using the backward method [20],[21]. The PR controller with the delay compensation and discrete PR controller is shown in Figure 3.



(a) The PR controller with the delay compensation (b) The resulting discrete PR controller

Figure 3. Block diagram of PR controllers.

2.1. Design the PR controller for the inner current loop

From Figure 1 and Figure 2, the plant transfer function of current control loop in series converter is determined as follows:

$$G_{iv}(s) = \frac{i_{sa}(s)}{v_{sa}(s)} = \frac{1}{N^2}\frac{sC}{s^2L_\sigma C + sRC + 1}. \quad (6)$$

The cross-over frequency is usually selected to be far lower than the sampling frequency f_s. On the other hand, the cross-over frequency f_c is significantly higher than the grid frequency f_1. From (3), $\left|G_{PR}(j\omega)\right|_{\omega=\omega_C} \approx k_{pc}$. The parameter k_{pc} of the PR regulator for the current loop is thus determined as follows:

$$\begin{aligned}&\left|G_{PRc}(j\omega)\right|_{\omega=\omega_{ci}}\left|G_{iv}(j\omega)\right|_{\omega=\omega_{ci}} = 1\\ &\rightarrow k_{pc} \approx \frac{1}{\left|G_{iv}(j\omega)\right|_{\omega=\omega_{ci}}}.\end{aligned} \quad (7)$$

Next, based on the desired phase margin PM_c of whole system, the parameter k_{rc} of the PR controller for the current loop is chosen such that:

$$PM_c = \angle G_{PRc}(j\omega)\big|_{\omega=\omega_{ci}} + \angle G_{iv}(j\omega)\big|_{\omega=\omega_{ci}} + 180^0. \tag{8}$$

Therefore, the parameter k_{rc} of the PR regulator is determined as follows:

$$\arctan\left[\frac{k_r \omega_{ci}}{k_p\left(\omega_1^2 - \omega_{ci}^2\right)}\right] = A_c \tag{8}$$

$$\Rightarrow k_{rc} = \frac{\tan(A_c)k_p\left(\omega_0^2 - \omega_{ci}^2\right)}{\omega_{ci}}, \tag{10}$$

where $A_c = PM_c - \left[G_{iv}(j\omega)\big|_{\omega=\omega_C} + 180^0\right].$

2.2. Design the PR controller for the outer voltage loop

From Figure 2, the plant transfer function of voltage control loop in series converter is determined as follows:

$$G_{vi}(s) = \frac{v_{inj}(s)}{i_c(s)} = \frac{1}{Cs}. \tag{11}$$

The magnitude-frequency and phase-frequency response of $G_{vi}(s)$ can be written as follows:

$$G_{vi}(j\omega) = \frac{1}{C\omega}, \qquad \angle G_{vi}(j\omega) = -90^0 \tag{12}$$

$$\begin{aligned} &\left|G_{PRv}(j\omega)\right|_{\omega=\omega_{cv}} \left|G_{iv}(j\omega)\right|_{\omega=\omega_{cv}} = 1 \\ &\Rightarrow k_{pv} \approx C\omega_{cv}. \end{aligned} \tag{13}$$

From the desired phase margin PM_v of whole system, the parameter k_{rv} of the PR controller is chosen such that:

$$PM_v = \angle G_{PRv}(j\omega)\big|_{\omega=\omega_{cv}} + \angle G_{vi}(j\omega)\big|_{\omega=\omega_{cv}} + 180^0. \tag{14}$$

Therefore, the parameter k_{rv} of the PR regulator is determined as follows:

$$\arctan\left[\frac{k_{rv} \omega_{cv}}{k_{pv}\left(\omega_1^2 - \omega_{cv}^2\right)}\right] = A_v \tag{15}$$

$$\Rightarrow k_{rv} = \frac{\tan(A_v)k_p\left(\omega_0^2 - \omega_{cv}^2\right)}{\omega_{cv}}, \tag{16}$$

where $A_v = PM_v - 90^0$.

3. RESULTS AND ANALYSIS

3.1. Simulation results

The proposed control topology is validated using MATLAB/Simulink/Simpower Systems. The DC voltage is 700VDC while the transformer turns ratio N is 2. The total of leakage inductance L_s in (1) is 0.2975mH, the filter capacitance C is 30µF, and the switching frequency f_s is 5kHz. Unipolar PWM technique is implemented to control the switching of the IGBT switches of the H-bridge inverters [19]. The phase margin and cross-over frequency of current loop are chosen to be 45^0 and 500Hz, respectively. The phase margin and cross-over frequency of voltage loop are chosen to be 45^0, and 200Hz, respectively.

This paper investigates the following transient scenarios of the grid voltage: single-phase 55% voltage sag (0.1-0.2s), double-phase 55% voltage sag (0.25-0.35s), and three-phase 70% voltage sag (0.4-0.5s). The RMS value of the phase-to-phase grid voltage is 380 VAC, i.e. the phase-to-neutral voltage is 220VAC. The three-phase voltage sags are created by a programmable voltage source in Matlab/Simpowwer Systems.

Figure 4 and Figure 5 show the response of the DVR system with the proposed control during the supply voltage sags. It is clear that the DVR is able to restore correctly to the nominal value within just one cycle of grid voltage (20 ms). The overshoots of the load voltage are negligible, and the steady-state error of injected voltage and inverter current is eliminated in the three cases in Figure 5. A dramatic part of the delay is due to the calculation of the root mean square (RMS) of supply voltage.

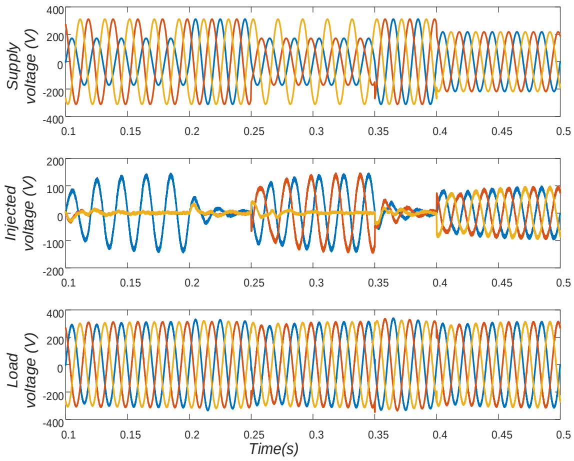


Figure 4. Measured response of DVR (grid voltages, voltages injected by the DVR, load voltages)

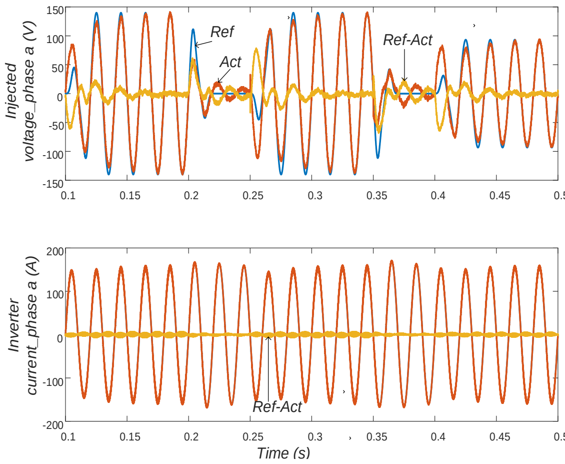


Figure 5. The reference and actual voltage and current of phase a

3.2. Hardware-in-the-loop experimental results

The proposed control is also verified in HIL environment using a Typhoon device. This device consists of an HIL402 card that simulates grid source, load, and three full-bridge with a common DC-link capacitor using IGBTs. The hardware system is simulated in real time on the HIL platform with a time step of 1 µs, which is close to the physical model. The carrier frequency of the PWM is 5 kHz. The voltage and current controllers as well as PLL are implemented in the DSP TMS320F2808 card.

All data of HIL is recorded by the Typhoon HIL Control Center Software and shown in Figure 6. The load voltage and the injected voltage from the DVR system are measured using the oscilloscope HAMEG –200MHz at test points in the HIL DSP interface of Typhoon. These voltages are shown in Figure 7. It is clear that the responses of the HIL experimental results are similar to those in the simulation results in MATLAB. The DVR system is able to regulate the load voltage with an ignorable overshoot within an acceptable period (less than 20 ms) in the injection mode. When supply voltage stable at the nominal value, the DVR system is operated in the standby mode, which means no voltage is injected.

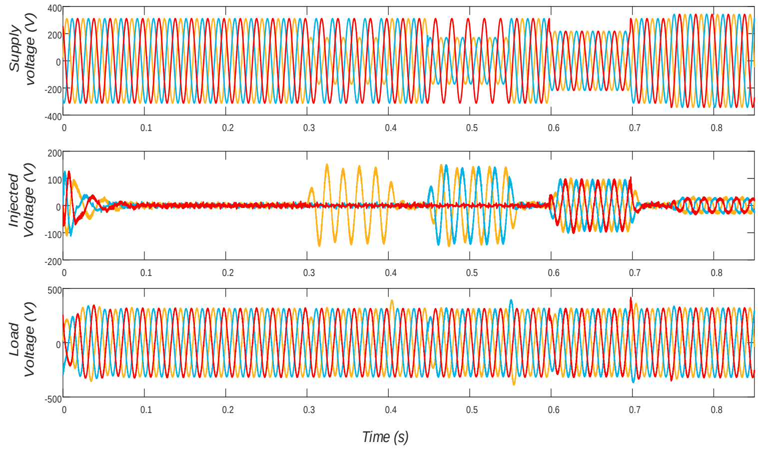
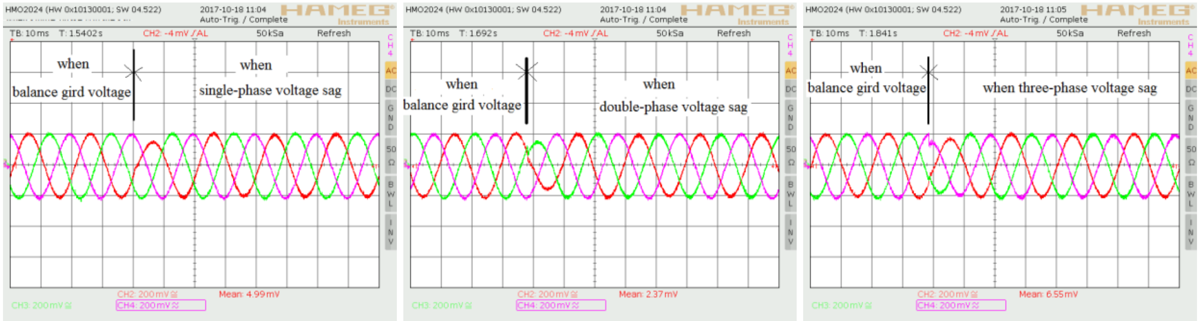
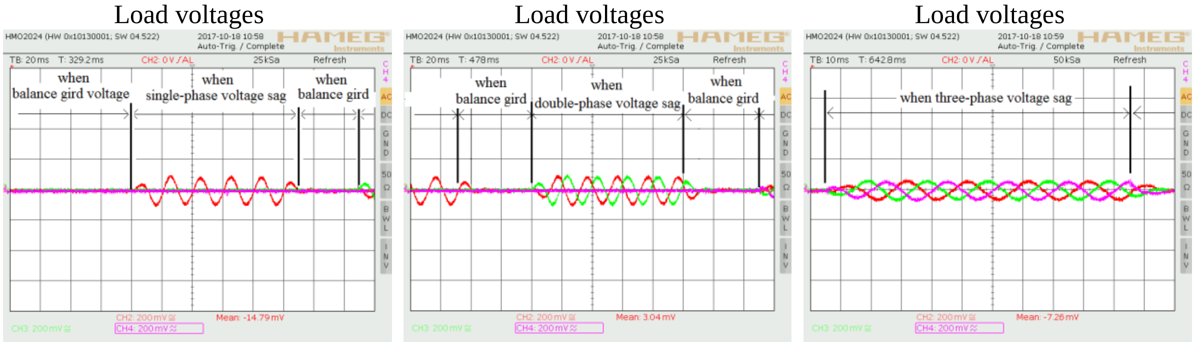


Figure 6. HIL experimental results: supply voltage, injected voltage, and load voltage



Load voltages | Load voltages | Load voltages



Voltages injected by the DVR. a. Single-phase 55% voltage sag | Voltages injected by the DVR. b. Double-phase 55% voltage sag | Voltages injected by the DVR. c. Three-phase 70% voltage sag.

Figure 7. HIL experimental results in different operating conditions

4. CONCLUSION

This paper proposes a systematic and generalized design method for PR controllers of three single-phase full-bridge inverters with a common DC-link capacitor in DVR systems. The proposed control is designed for each single-phase inverter, taken into account the model of the series transformers. MATLAB and the HIL experimental results validate the performance of voltage in scenarios: single-phase and double phase voltage sags up to 55%, and three-phase voltage sag up to 70% within acceptable periods. The results prove that the DVR system is able to protect the load from voltage sags due to these various types of faults. Such promising results create a crucial foundation for the application of DVR systems in industry.

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